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BULLETIN NO. 38

BY Arthur E. Creech USO

F I F T H R E P O R T
September 1952

A Report on

MILLIMETER WAVE RESEARCH

Contract Nonr-687(00)

Prepared by Bell Telephone Laboratories, Inc.

On Behalf of

Western Electric Company, Inc.

Contains intelligence produced outside of the work performed under contract Nonr-687(00)

This work has been supported in part by the Air Force, Army Signal Corps and the Office of Naval Research.

Edited by

S. E. Miller
J. R. Pierce

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SIGNAL CORPS ENGINEERING LABORATORIES
EVANS SIGNAL LABORATORY
BELMAR, NEW JERSEY

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Fifth Quarterly Report
September 1952

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1. Components and Measuring Techniques

1.1 Measuring Circuits (D. H. Ring)

The 5.4 mm component construction program is substantially completed, and the initial requirements for several bench setups and a propagation test circuit are either on hand or have been delivered to those interested.

A problem has arisen in connection with attenuator resistor cards. We have been using 0.002 inch mica strips coated with a resistance paint. Crude dipping and spraying techniques do not yield sufficiently reproducible results to be satisfactory. In addition, the attenuation appears to change with age over a period of several weeks or months. Stable evaporated metal films can be reproduced very accurately, and appear to offer a good solution to this problem. A number of mica strips have been sent out for coating and will be used to investigate the properties of this type of attenuator element.

1.2 Noise Source (W. W. Mumford)

Our first thoughts concerning a gaseous discharge noise source for use in the millimeter wavelength region were focused on adapting a 24,000 mc noise source by means of a tapered section of waveguide. Since satisfactory 24,000 mc noise sources are available and on hand, this seemed to be a handy expedient to adopt. However, we have recently had some discouraging experience along similar lines, in which a 9,000 mc noise source was tried at 24,000 mc and found to have a lower effective temperature at the higher frequency. A tentative explanation for this is that the conductance seen looking into the circuit is not due entirely to the plasma in the positive column, but partly due to losses in the envelope and the connecting leads of the lamp. These losses are in the cool portion of the lamp and hence reduce the effective temperature. They show up at the higher frequency and not at the lower frequency because the portions of the lamp outside the main waveguide are housed in metal shields, which form branch waveguides that are beyond cutoff at the lower frequency and not at the higher frequency. Because of this experience, we feel that it is important to keep the discharge tube diameter small enough to form cutoff waveguides at the operating frequency.

Accordingly, a small diameter discharge tube has been made and tested. Its noise temperature, measured at 6.2 mm wavelength by Mr. W. M. Sharpless, is 10,650° Kelvin, which makes it useful for measuring noise figures of less than, say, 25 or 30 db.

The active portion of the lamp is 2" long and consists of a fused quartz tube having an outside diameter of .075" and an inside diameter of .055", filled with argon at a pressure of 25 mm Hg. These diameters were dictated by the requirement that the shielded portions should be beyond cutoff.

The circuit in which the tube is mounted consists of a rectangular waveguide with a 180° E plane bend of 1-1/4" radius. The tube is mounted so as to intersect the bent waveguide tangentially. Although it is easy to fabricate this type of circuit, the electrical performance is not too satisfactory. The return loss of the tube and circuit, when the far end is terminated in the characteristic impedance of the waveguide, is only 12 db with 20 ma discharge current and 4 db with no current. A better circuit is now being built, in which the lamp enters a straight waveguide at an angle of 10°.

1.3 Millimeter Wave Pistons (W. W. Mumford)

Some experimental work has been done on non-contacting pistons at 4000 mc for a scaling to 5 mm. These pistons consist of a pair of non-contacting rectangular slugs separated by an appropriate distance and supported at one end by a sliding slug which may or may not make contact all around. Working in 1.872" x .872" waveguide at 3700 to 4500 mc, it is seen that the best length for the non-contacting slug is a quarter of the air wavelength, and that the best separation is about a sixth of the guide wavelength. These results apply to slugs 1.862" x .862" in cross section connected by a strip 1.862" x .125". Two such slugs gave over 30 db attenuation over the band, and, when supported by the sliding member to form a piston, had a satisfactory return loss of less than .07 db over the band, with less than .08 Z_0 reactive component. Wobbling this piston did not alter these values appreciably. Models scaled down to the millimeter waveguide size are now being fabricated.

2. Crystal Conversion Loss and Noise Figure Measurements (W. M. Sharpless)

Continuing with the investigation of conversion losses of silicon rectifiers and noise figure improvements, several important advances have been made since the last quarterly report.

A balanced converter for use as a first detector in our millimeter wave receivers was designed, and one unit has been assembled and tested at a wavelength of 6.3 mm. The unit employs two silicon rectifiers, as nearly identical as possible, fed in a balanced hybrid-tee waveguide arrangement; the 60 megacycle output is taken by means of a balanced to unbalanced 20 megacycle wide I.F. transformer having a 72 ohm output impedance. Inasmuch as the same hybrid tee is used with one arm terminated for the unbalanced detector measurements, the effect of adding an additional rectifier is to increase the signal output level 3 db (in the balanced detector case). The noise level at the receiver output with the balanced detector was measured to be 3.2 db lower than with the unbalanced detector. (The I.F. amplifier has a noise figure of 5.7 db). Thus, under these conditions, an improvement of slightly over 6 db in signal to noise was realized by the use of balanced detection. This noise figure improvement was further verified by making an over-all noise figure measurement on the complete receiver, making use of Mr. Mumford's new millimeter wave noise source (reported on separately by Mr. Mumford). The use of this noise source also confirms other measurements of the over-all receiver noise figures arrived at earlier by separately measuring the crystal conversion loss, the I.F. amplifier noise figure and the Y factor.

There has been some evidence that a portion of the measured conversion loss of our silicon rectifiers is due to the losses in the Micarta bead supporting the point contact terminal of the rectifier. With this in mind, a new terminal section was designed, which eliminated this section of Micarta and replaced it with a small low-loss by-pass condenser arrangement. This condenser consists of a quarter wavelength long, low impedance, coaxial section assembled and insulated by Araldite cement. Ten of these new type rectifier cartridges were made and measured for conversion loss at 6.3 mm. It was found that the average conversion loss of this new group was 8.6 db, whereas the best rectifier we had before was 8.5 db. The best rectifier obtained from this group had a conversion loss of 7.5 db. Large scale production of this new unit might, however, prove difficult.

A new receiver was recently put into operation at 5.4 mm, and we now have in operation complete measuring receivers for both 5.4 and 6.3 mm. In connection with the setting up of this new receiver, the power output of one of our harmonic producers at a wavelength of 5.4 mm was measured. The primary oscillators used to drive the harmonic producer for this wavelength are the Raytheon QK289 tubes. Using eight different driver tubes, the harmonic producer output power was measured by the use of the calorimeter. It was found that on an average the power delivered at 5.4 mm was very nearly 1.5 milliwatts. The power delivered for the case of the best driver tube was 4.0 milliwatts, while with the poorest driver tube only 0.25 milliwatts was realized.

Preliminary conversion loss measurements have now been made at a wavelength of 5.4 mm, and the difference in conversion loss of a silicon rectifier at 5.4 and 6.3 mm has been measured. On the basis of a few measurements made on a single rectifier at both wavelengths, it was found that the conversion loss at 5.4 mm was 0.8 db greater than at 6.3 mm. (The same BO crystal current was used in each case.) It is unlikely that this difference will be the same for all rectifiers, and it is the plan therefore to measure the conversion losses at both wavelengths on a group of rectifiers from which a more reliable difference figure will be determined. The results of these tests will be given in the next report.

A portion of this work has been carried out using Bell System funds as part of the Radio Research Program.

3. Dielectric Waveguides (A. G. Fox)

Experiments with dielectric waveguide directional couplers have been continued, and good results have at last been obtained using accurately dimensioned rods which have been shaved to size by pulling through hardened steel dies. In the course of this work it was discovered that maximum power transfer from one guide to the other was greater when both guides shared the curvature necessary to couple them together, rather than when one was straight and the other curved. Thus, best coupler performance is obtained when the coupler is symmetrical.

Experiments are in progress on a double polarization type of coupler suggested by S. E. Miller, comprising a main line of circular cross section rod which can propagate two polarizations of the dominant wave, and a coupled line of rectangular cross section which is designed to tap off 100% of the power of one of the two possible polarizations on the round rod. Due to the difference in phase velocities for the two possible polarizations on the rectangular rod, the other polarization can not be coupled off of the circular rod. Several pairs of round and rectangular dielectric rods have been prepared which couple power satisfactorily for the desired polarization. It appears that when the phase velocities for a pair of these rods have been made equal, then the attenuations for the two rods are also approximately equal.

A section of .062" x .188" polystyrene rod was mounted between launching horns so that the electric vector was parallel with the wide side. For this polarization of wave, the field is much more tightly confined to the rod and it was observed to have much greater flexibility than when E is parallel to the short side. Thus, a 180° bend could be made in a particular patch cord two feet long with a resulting loss in transmitted power of no more than 1/10 of a db, where the same length of rod, when excited with a polarization parallel to the short side, produced a loss of nearly 4 db for a 180° bend. Notwithstanding the improved flexibility, the increase in total attenuation due to the use of this more closely bound wave was only about 1.5 db. Thus for flexible patch cords the more tightly bound polarization is probably to be preferred.

4. Flexible Waveguides for TE_{01} Waves

4.1 Moderate-Loss Structures (A. P. King)

As indicated in the preceding quarterly report, some trouble has been encountered with the earlier samples of spaced-ring line in maintaining accurate coaxial alignment of the rings. This difficulty is attributed to the poor physical properties of the L P rubber which was employed as the flexible bonding jacket.

In substituting a rubber more suitable for the ring line structure, the use of neoprene appears very attractive. In addition to the excellent physical stability possessed by neoprene, it is also very stable chemically and can be loaded with conductive material to vary the resistivity of the rubber over a wide range; at least $10^3 - 10^9$ ohms/cm^3. The latter characteristic provides a wide range of control in the amount of R. F. dissipation in the radial line sections.

A change in the design has also been effected to facilitate the fabrication of spaced-ring lines. As illustrated in Fig. 1, the new structure comprises an ensemblage of copper and neoprene washers. Before assembly the neoprene washers were coated with a dry back adhesive whose bond could be activated by the application of heat. The spaced-ring line was assembled by stacking the copper and neoprene washers alternately on a mandrel and then heating the entire structure to 250° F for a few minutes to provide a copper-to-rubber and rubber-to-rubber bond. After the structure cooled, the mandrel was removed, leaving the completed spaced-ring line. It is to be noted that a line constructed

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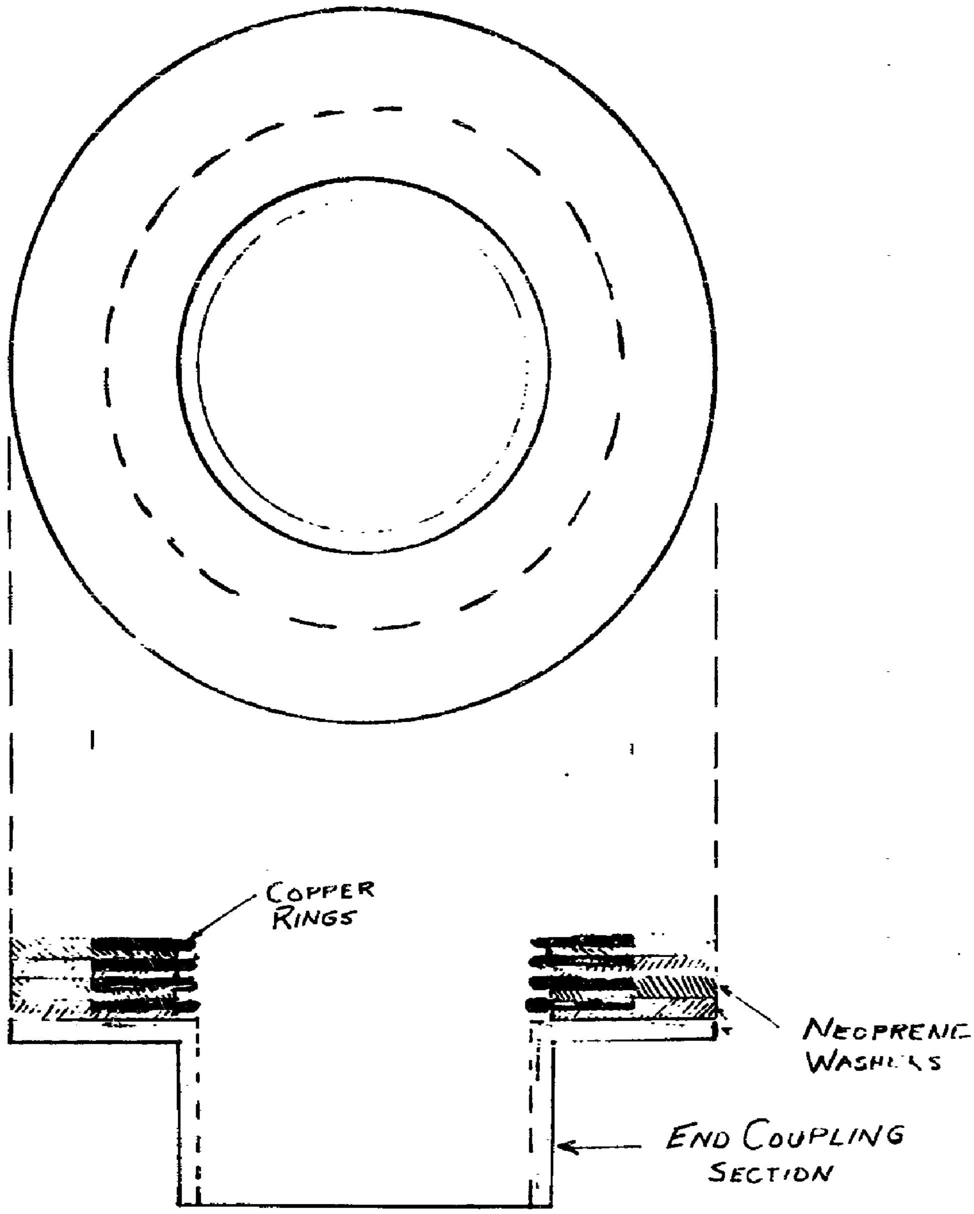

FIG. 1

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ENG. APKing	SPACED-RING LINE
	SCALE NONE
	BELL TELEPHONE LABORATORIES INCORPORATED, NEW YORK
	NO. OF SHEETS PER SET — SEE SHEET 1
APP'D / CH. / DR.	SHEET

ISSUE

in this manner does not require spacing rings which have to be dissolved out after the mandrel has been removed.

A few short samples which employ this alternate washer arrangement have been built for physical tests and the results appear most promising. As soon as a sufficient quantity of washers become available, longer lines will be constructed to provide samples for electrical tests.

A program has been started for the construction and study of spaced-ring lines of a smaller diameter. Two sizes have been chosen, .4375" and .875" inside diameter, for applications in the 5.0 to 6.0 mm range. The design and construction of associated waveguide components in these two sizes are also under way.

4.2 Low Loss Structures (M. Aronoff)

During the past year, a program for obtaining and evaluating preferred structures (structures other than cylindrical round copper tubing) for propagating the TE_{01} wave has been carried on for Bell System applications. As part of this program the properties of a spaced-ring circular electric waveguiding structure were investigated.

One such study consisted in measuring a demountable spaced-ring structure at about 9000 mc. The spaced-ring waveguide was built up of a series of flat copper plates, 5" square, punched to provide a 4.732" diameter center hole and four 3/8" corner holes. Both 1/16" and 1/32" thick plates were used. The plates were assembled on four 3/8" studs and were spaced apart by 5/8" O. D., 3/8" I. D. washers, which fit over the studs. The structure was assembled in 2' lengths on an accurately ground straight mandrel. End plates 1/4" thick and soldered to a section of 5" O.D., 4.732" I. D. solid copper tubing were provided to get a transition from the spaced-ring waveguide to 5" solid round waveguide. The entire structure was held together by longitudinal pressure exerted by nuts at the ends of the studs and bearing on the end-plates. Overall straightness and concentricity tolerances of 2 mils were achieved.

The radial depth of the spaces between the rings varied around the circumference. The minimum depth was 1/8" - which was twice the width of the air gap for the widest spacing used. Bearing in mind that the attenuation of a waveguide beyond is 27.3 db per guide width for an infinite length guide, and that the circular electric field is zero at the wall, it can be seen that the circular electric wave leakage through the radial gaps is very small for such a structure. For modes with longitudinal wall currents, however, the radial waveguide propagates energy, and so the undesired modes leak energy to the radial lines and thus are attenuated.

As can be seen in Fig. 3, the spaced-ring line samples constructed by this method do not provide an energy absorbing envelope for the unwanted modes, but do have the advantage of relatively easy construction so that the parameters may be varied over a reasonable range in a reasonable amount of time.

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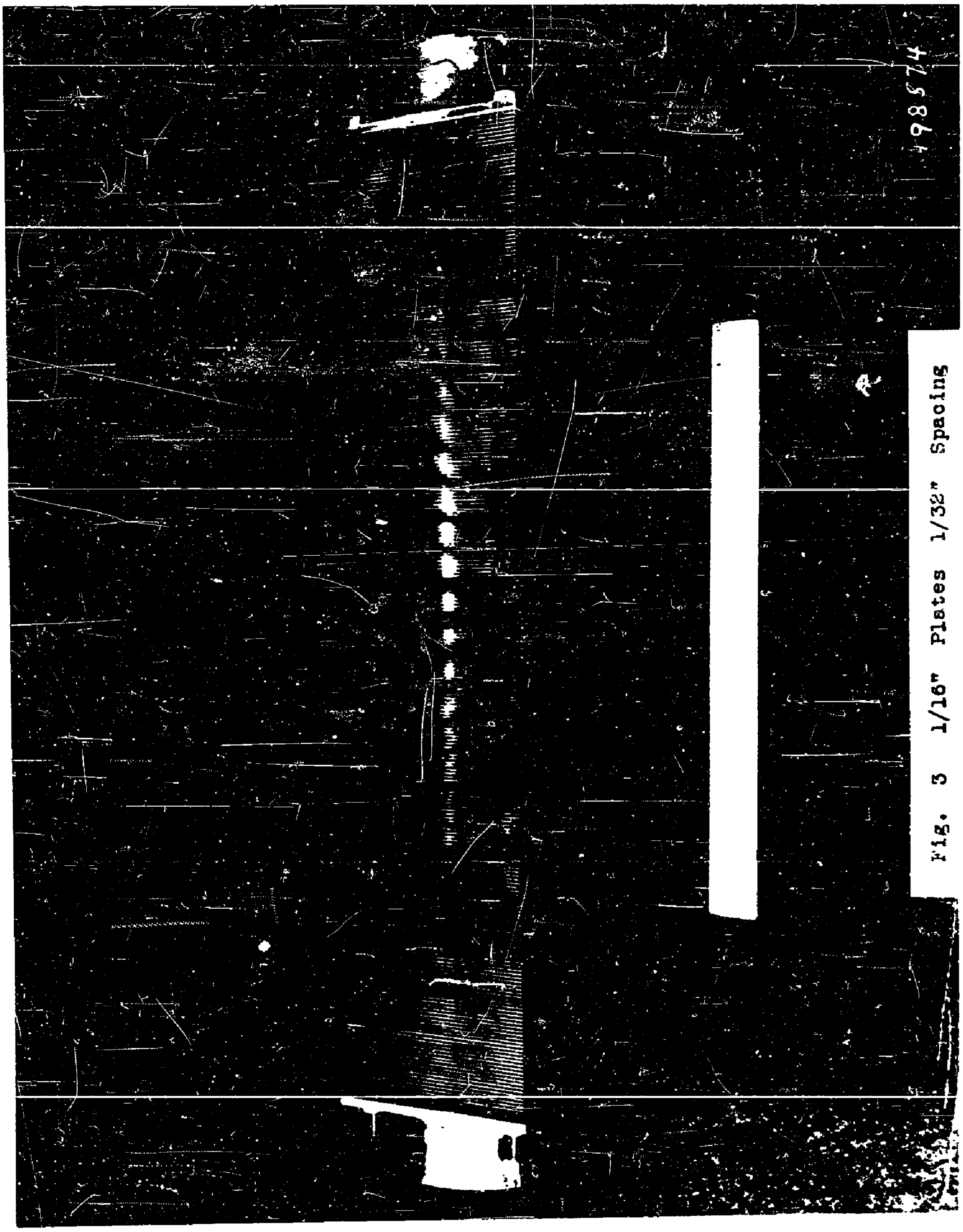

Fig. 3 1/16" Plates 1/32" Spacing

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The attenuation of several samples of spaced-ring line (similar to the sample shown in Fig. 3) was measured for the TE_{01} mode as well as for several modes having radial electric field components.

As has been shown by W. A. Tyrrell and B. C. Wood of the Bell Telephone Laboratories, the circular electric mode attenuation of short lengths (a few feet) of 5" pipe can be measured with reasonable accuracies at 9,000 mc (better than 5% if sufficient time is taken for the measurement) by the pulse decrement method. This method consists essentially in measuring the rate of decay of a pulse of energy introduced into a cavity, the wall of which is the waveguide being measured. Attenuations as small as a few db per mile have been accurately measured by this method at 9,000 mc with pipe samples 8 to 20 feet in length on a laboratory bench setup. The equivalent loaded cavity Q's are of the order of 6 x 10^{-5}. This method was used to obtain the TE_{01} loss of the spaced-ring samples.

To measure the mode filtering characteristic of the spaced-ring structure, the radial probe technique described by M. Aronoff,* was used. Pure modes (TE_{11}, TM_{01}, TE_{21}, TM_{11} and TE_{31}) were generated, and the measurements were made by comparing the radial probe output (magnitude and pattern) with and without the spaced-ring sample in the system. Measurements were made as

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*M. Aronoff, "Radial Probe Measurements of Mode Conversion in Large Round Waveguide with TE_{01} Mode Excitation" (submitted to Proceedings I.R.E.)

a function of sample length for $\frac{s}{t} = 1$ and for two values of $\frac{s}{t}$ for a fixed length of 60.5 cm where $\frac{s}{t}$ is the ratio of ring spacing to ring thickness. The attached Fig. 2 illustrates the measurement method and gives the results.

The measurement results show, as is indicated in the following tables I and II and Figs. 2, 4, and 5, that, under the conditions of measurement, the attenuation of a close packed spaced-ring line (zero spacing between plates) to the TE_{01} mode is within about 35% of the theoretically attainable attenuation for geometrically perfect tubing of the same d.c. conductivity. As the plates are spaced apart, the TE_{01} attenuation increases, rapidly at first and then more gradually until it approaches a value of a little less than four times (in db/unit length) the TE_{01} attenuation of perfect tubing. (This increase appears to be asymptotic.) At the same time, the ratio $\frac{\text{loss to undesired radial E modes}}{\text{loss to desired modes}}$ increases to a value of at least about 2×10^3 for the lowest loss undesired radial E mode we could measure.

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ISSUE 1 5-26-52 M ARONOFF S.A DARBY

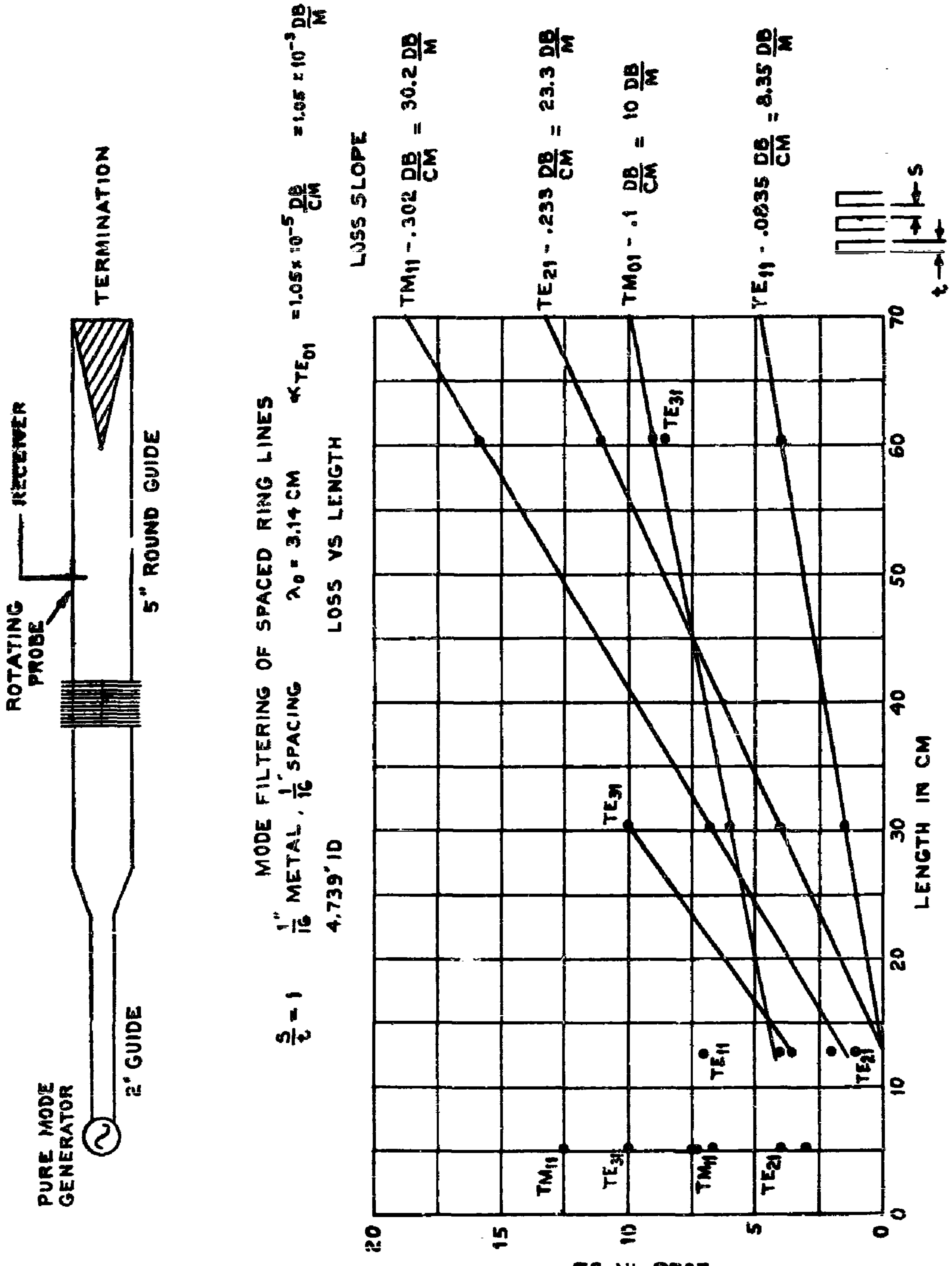

FIG 2 - RADIAL PROBE MEASURING SET UP AND MEASURED RESULTS

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ISSUE 1, 8-29-52 M. ARONOFF - S.A.D.

FIG 4 - SUMMARY OF SPACED RING LINE MEASUREMENT

FIG 4a - $\frac{\alpha \text{ EXPERIMENTAL}}{\alpha \text{ THEORETICAL}}$ VS $\frac{s}{t}$ FOR RING WAVEGUIDE

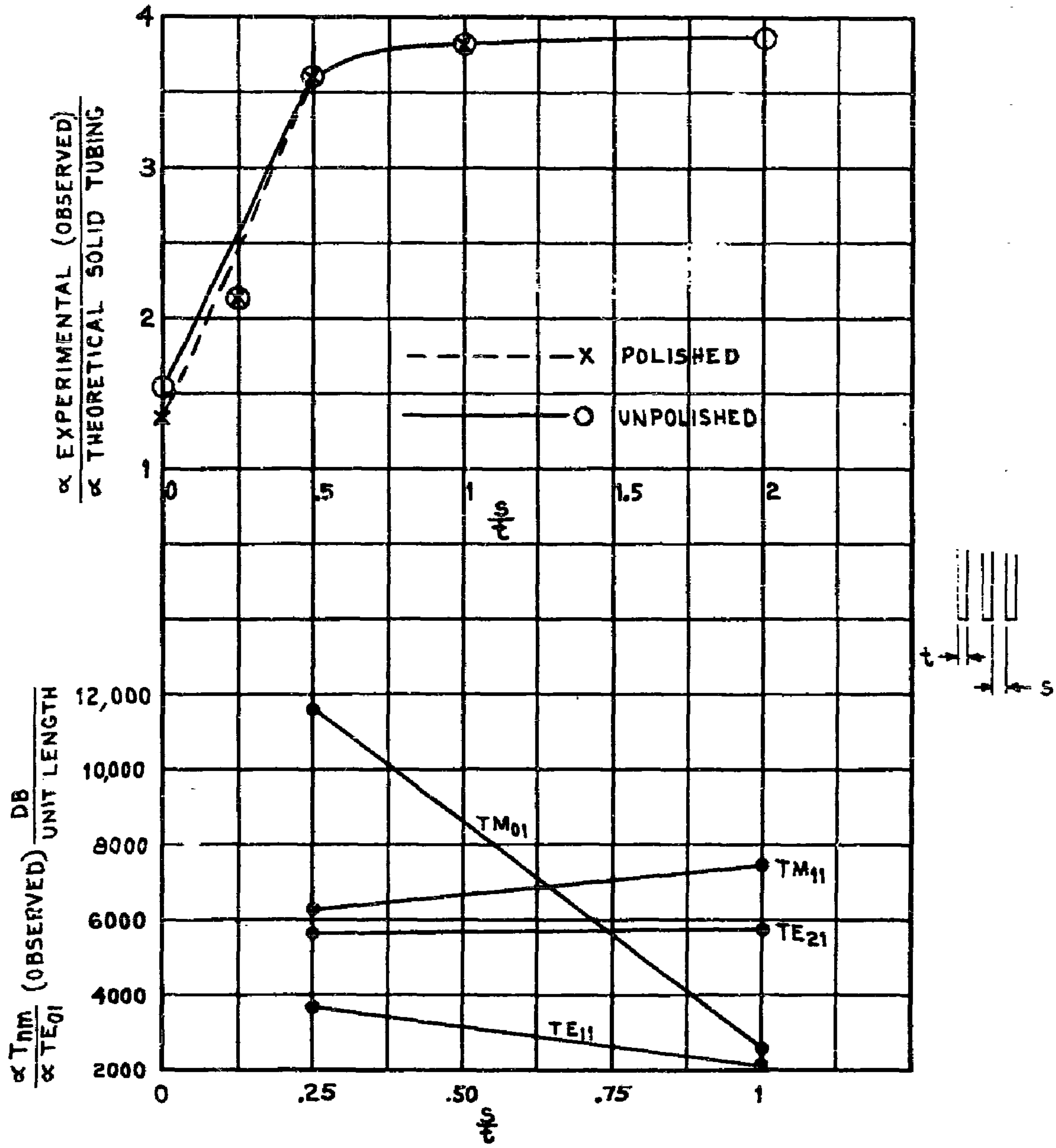

FIG 4b - $\frac{\alpha T_{nm}}{\alpha TE_{01}}$ (OBSERVED) $\frac{DB}{\text{UNIT LENGTH}}$ VS $\frac{s}{t}$ FOR SPACED RING LINE

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MODE FILTERING OF SPACED RING LINES

DB LOSS VS $\frac{S}{t}$ $\lambda_0 = 3.14$ CM

60.5 CM LENGTH

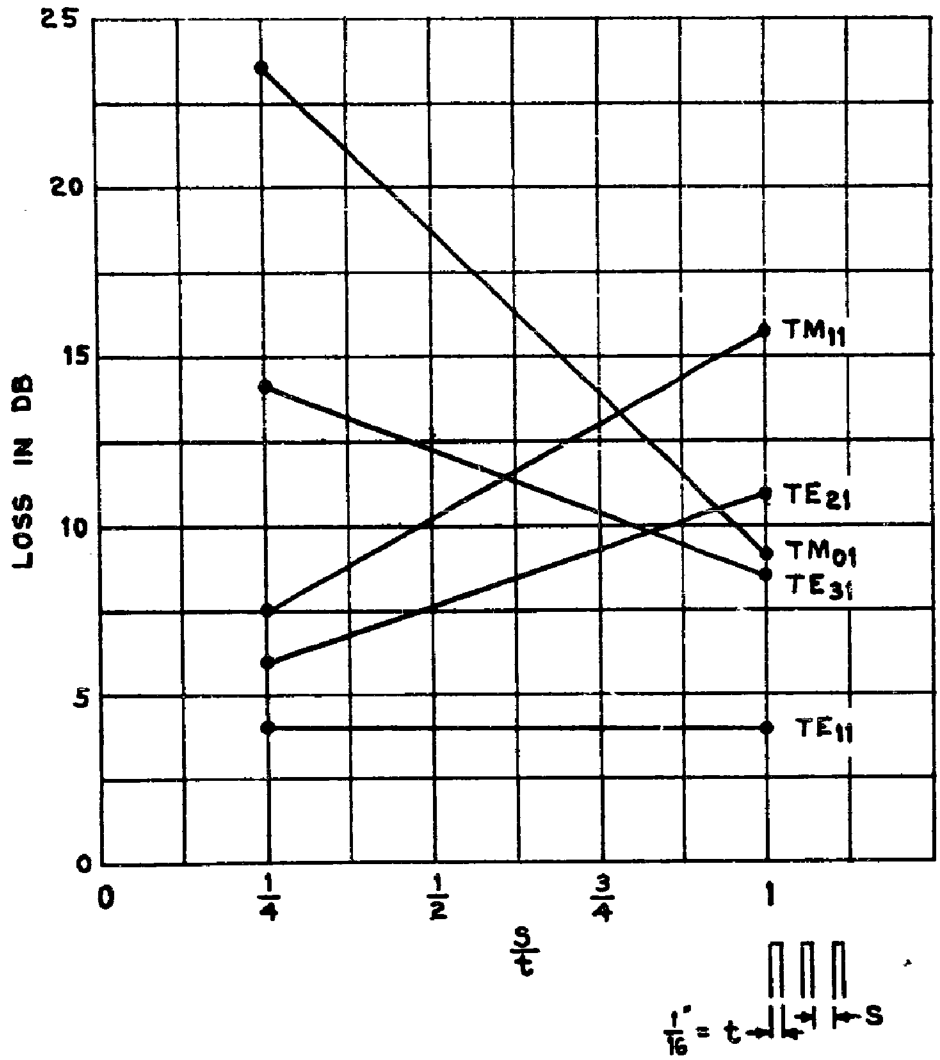

FIG 5 - MODE FILTERING MEASUREMENTS

5-26-52

ISSUE 1

S.A. DARBY

M. ARONOFF

APP'D

SCALE

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A

Table I - Unpolished Plates

Plate Thickness - t	Plate Spacing - s	$\frac{s}{t}$	$\frac{\alpha \text{ exp.}}{\alpha \text{ theor.}}$
1/16"	0	0	1.55
	$\frac{1"}{64}$	$\frac{1}{4}$	2.13
	$\frac{1"}{32}$	$\frac{1}{2}$	3.65
	$\frac{1"}{16}$	1	3.79
1/32"	$\frac{1"}{64}$	$\frac{1}{2}$	3.64
	$\frac{1"}{32}$	1	3.81
	$\frac{1"}{16}$	2	3.84

The above measured results are the smoothed average of both single 2' lengths of spaced-ring line and two 2' lengths joined in tandem.

For a four foot length of commercial copper tubing, we measured $\frac{\alpha \text{ exp.}}{\alpha \text{ theor.}}$ to be 1.28. Tyrrell, in a much more careful determination of carefully polished commercial copper tubing,has obtained $\frac{\alpha \text{ exp.}}{\alpha \text{ theor.}}$ as low as 1.20.

Table II Polished Plates

Plate Thickness - t	Plate Spacing - s	$\frac{s}{t}$	$\frac{\alpha \text{ exp.}}{\alpha \text{ theor.}}$
	0	0	1.36
$\frac{1"}{16}$	$\frac{1"}{64}$	$\frac{1}{4}$	2.13
	$\frac{1"}{32}$	$\frac{1}{2}$	3.66

The work described above was carried out using Bell System funds exclusively, as part of the Radio Research Program.

5. Point Contact Rectifiers

5.1 Cylindrical Cartridge Units (R. S. Ohl)

Studies of the mechanical properties of the tungsten point contact spring have been made. Microscopic measurements show that the contact area is proportional to the contacting force. The constant for Westinghouse S-0-13 tungsten wire was found to be 8×10^{-8} sq. cm. per gram of force.

It was found possible to calculate the force on the point from the dimensions of the spring configuration and the measured mechanical constants of the wire. Experimental results confirmed the calculated values of force. The force was found to be proportional to the axial displacement of the point; therefore the contact area can be estimated from the axial displacement of the point when setting up the contact.

The maximum current which can be drawn from a point per gram of contact force, without depreciation of the rectification characteristic, has been found to be 2 ma average current, corresponding to about 9 ma of peak current for doped silicon bombarded with 30 KV helium ions. This was tested up to 104 grams of force.

A study of the electrostatic field in our bombarding apparatus resulted in information which indicated that an improvement in the uniformity of bombardment might result from a slight addition to one of the electrodes. After this alteration had been made, tests showed a considerable improvement in the uniformity of ion dosage.

Calculations of the ratio of displacement currents to the conduction currents indicate that with the silicon we are now using, having been doped with .02% boron, the conduction currents far exceed the displacement currents. It was assumed that the high frequency resistivity remains substantially the same as that measured at low frequencies. This indicates that there is no electrical advantage in cutting the silicon thinner and this appears to be supported by the results of a few exploratory experiments.

Instability and some other unaccounted for variable effects in millimeter rectifiers have been found to be due to loose pins in the insulating support at the center conductor. This condition set in when the moulding procedure was changed to center the pins more precisely. The difficulty seems to be due to a trapping of gas around the pins during moulding. A few of these moulded parts have been made up with the mould vented. These mouldings appear to be satisfactory. The loose pin situation has been found to be more detrimental to mixer type units than to the harmonic type or the signal detection type. In these latter types the spring seems to take up the displacement of the loose pins; however, in mixer types no pin movement can be tolerated. This difficulty is now being corrected.

While investigating the source of our difficulties with loose pins in our bakelite mouldings, mechanical strength tests were made. The insulation beads were found to withstand a crushing pressure of 38,000 p.s.i. The pins in a tight moulding job could be dislodged by a force of about 100 pounds.

Some temperature tests were run on cartridge type rectifiers. The rectifiers showed no permanent change in properties even though the temperature reached 210° C. This indicated that the bakelite resin type of insulation is satisfactory up to this temperature. In order to make this test possible, the silicon was pressed into its coin silver support with a force of one and a half tons. The usual method of fastening the silicon is by soldering it in place, but obviously at the temperature of this test the solder would have melted.

Some test mixers were made up with a pin holding arrangement designed and made up by W. M. Sharpless. This arrangement has a by-pass capacity about 10 times greater than the capacity in the mixers currently in use. The conversion loss of mixers with this higher by-pass capacity is about 1 or 2 db lower than with the older arrangement. (See the report of Mr. Sharpless.) This situation is being actively followed up.

Millimeter rectifier units are currently being furnished to members of other departments working on this project.

A portion of this work has been carried out using Bell System funds as part of the Radio Research Program.

5.2 New Type Crystal Rectifier (A. B. Crawford)

Work is continuing on the crystal holder, in which the rectifier is mounted directly in the waveguide, to overcome the problem of mechanical instability in the by-pass condenser.

6. Millimeter Wave Propagation (A. B. Crawford)

Since components for use in the 5.4 mm band are now becoming available, it has been decided to attempt propagation observations on the two mile overland path between the Holmdel Laboratory and Crawford Hill in the 5.4 mm band rather than at 6.3 mm as originally planned. Preparation of the terminal locations, including building power and necessary tree cutting, is complete and most of the 5.4 mm equipment is at hand. The major difficulty is the inability to devote the time to this project that it deserves.

7. Faraday Rotation (A. G. Fox, M. T. Weiss)

Equipment has been set up to make Faraday rotation measurements in the range from 8.2 to 12.4 kmc. This wide band is necessary in order to determine the variations in loss and in rotation with frequency, as well as to aid in understanding the circuit problems involved in using ferrites in waveguides at millimeter wave frequencies.

Measurements made with this equipment indicate that the Faraday rotation, observed with a waveguide partially filled with a pencil of ferrite, increases with increase of frequency. The value of this increase depends on the diameter of the guide and the diameter of the ferrite pencil relative to a wavelength. The reason for this increase is apparent when one considers that the higher the frequency, the more closely bound is the wave to the ferrite and, therefore, the higher the rotation. For a waveguide completely filled with ferrite, a slight increase of rotation with frequency still remains, as shown by H. Suhl and L. R. Walker in Phys. Rev. 86, 122 L.

In making measurements of loss and reflection coefficient of a well tapered ferrite as a function of magnetic field, peculiar violent variations of these quantities were observed at certain frequencies. It is believed that these variations can be attributed to a partial conversion of the dominant TE_{11} mode to the TM_{11} mode by the tapered ferrite, which was of sufficient cross section to be able to maintain this higher order mode. As the magnetic field is varied, the effective electrical length of the ferrite changes so that the interference between the TM_{11} and TE_{11} modes can also vary.

In order to check the above hypothesis, experiments were performed using circularly polarized waves. As is well known, theory predicts that for a positively rotating circularly polarized wave, the permeability, μ, of the ferrite starts at unity, (at zero magnetic field), and then decreases, passes through zero and goes through resonance as the applied longitudinal magnetic field is increased. For a negatively rotating circularly polarized wave, on the other hand, μ starts at unity and then increases until saturation is reached, and finally approaches unity again at very high magnetic fields. From the above considerations it is evident that the negatively rotating wave should be more readily subject to mode conversion than the positively rotating wave, since the former has a µ greater than unity, while the latter has a µ which drops sharply below unity with applied magnetic field. Our experiments with circularly polarized waves confirm the above ideas since the negatively rotating wave showed many loss peaks, while the positively rotating wave had a smooth loss curve at fields below the resonance field value.

In the previous progress report it was stated that many ferrites having high loss at 9 kmc with zero applied magnetic field can have this loss substantially reduced by the application of moderate magnetic fields. When examined using circularly polarized waves it was found that this low-field loss has a peak not at zero field, but at rather small field values, and not for a positive circularly polarized wave but rather for a negative circularly polarized wave. Experiments with a wide variety of ferrites give substantially the above results with only a very few exceptions. No explanation for this near zero field loss has yet been formulated.

These circularly polarized wave experiments also revealed the surprising fact that for large diameter samples a resonant loss peak or peaks-occur not only at the resonant field value for the positively rotating wave, but also for approximately the same field value for a negatively rotating wave as well. It is believed that this effect is connected with the fact that for a waveguide propagating a circularly polarized wave of a given mode, the electric and magnetic field vectors are in general circularly polarized in the same sense only at the center of the waveguide cross section. Near the walls of the guide the transverse component of the rf magnetic vector is linearly polarized for the TE_{11} mode. For the TM_{11} mode there are not only regions where the transverse rf magnetic vector is linearly polarized but there are some where it is circularly polarized in the opposite sense from that at the center of the waveguide. No definitive experiments have yet been done to check this hypothesis.

Substantial evidence has been obtained which shows that at 9 kmc the loss for powdered and moulded ferrites is higher than for solid ferrites. This can be explained on the basis of the broadening of resonance in the moulded samples due to the wide variety of demagnetizing factors in the individual particles and perhaps also due to the inclusion of single domain particles. At 20 kmc and higher, the above increase in loss for moulded ferrites should disappear since even a broadened resonance would not have much effect at moderate fields. This latter prediction has not yet been checked.

By placing a ferrite Faraday element between two rectangular waveguides, it is possible to make an electrically controlled variable attenuator, when the ferrite is operated below saturation. Theoretical calculations show that this attenuator, when operated far from ferromagnetic resonance, should show no change in phase shift as the attenuation is varied. T. S. Benedict built such a device for operation at 24 kmc and found it to operate as predicted.

8. Millimeter Wave Electronics (S. D. Robertson, A. Karp)

8.1 Helix Traveling-Wave Tube

The remaining details of the assembly of the demountable model of this tube were worked out during this period. Special jigs and welding electrodes were devised by Mr. F. A. Braun for forming the ends of the helix and connecting them to the points of the central conductors of the coaxial input and output lines. Several practice connections were made in this way.

Difficulties were encountered in making vacuum seals in the coaxial lines with the tapered quartz beads previously described. The slightest mechanical stress caused chipping of the quartz at the thin edge of the taper. Tapered beads of low vapor-pressure thermosetting plastics, cast in place, were also tried without success. It has been finally found, however, that beads machined of "Araldite" CN501 and cemented in place with clear glyptal are satisfactory. It is expected that the tube will soon be ready for testing.

8.2 Spatial Harmonic Tubes

8.21 Circuit Structures

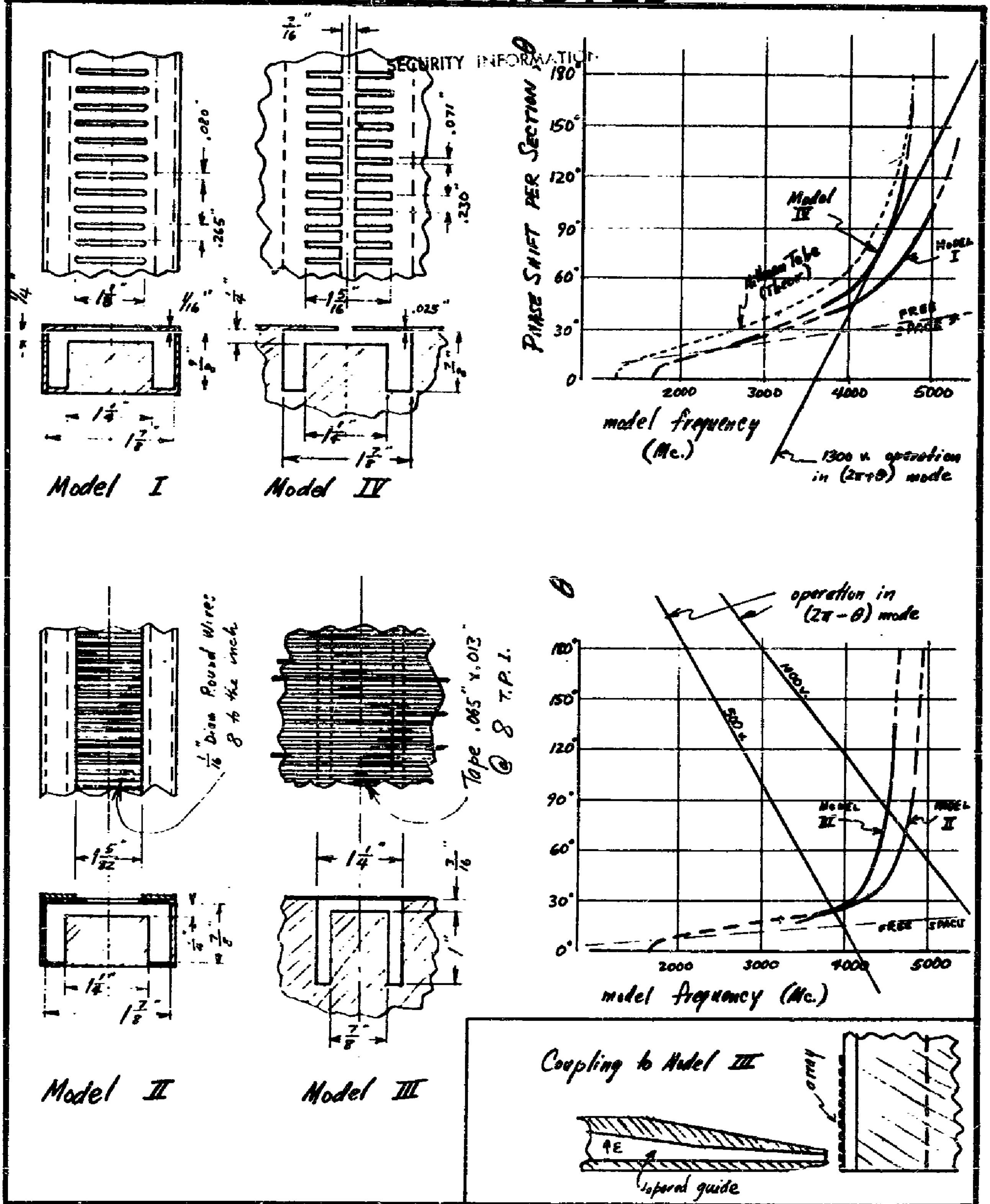
Work continued on the building and measuring of slow-wave circuit models based on the principle of a ridged rectangular waveguide whose broad face is perforated with iterated transverse resonant slots, the goal being to find structures with the best electrical characteristics consistent with a high order of simplicity of construction at millimeter wavelengths.

Models were tested at 4000-5000 mc, where the wavelength of the slow waves was determined by means of a moving probe or reflector. (Still slower waves, corresponding to spatial harmonics, were also observed.) Four typical models are sketched in Fig. 6, giving the observed phase shift per section as a function of frequency. Models I, II and IV were excited by "tapering out" the slots and the ridge so that the circuit could be joined directly to a rectangular guide. In these cases the input match was quite good, and insertion losses could be measured. To couple to Model III, it was found that a waveguide, that had been tapered down to a narrow E dimension, brought close to the next to the last wire as shown in Fig. 6 (bot.), would do the job fairly well. The merits of this coupling scheme are simplicity, compactness, and economy of space rather than optimum electrical characteristics.

8.22 New Tubes

8.221 Amplifiers

Model IV has a phase characteristic very close to that of the Millman tube, and the possibilities of duplicating the latter's performance with a tube utilizing this new struc-



SLOW-WAVE STRUCTURES for SPATIAL HARMONIC TUBES

Fig. 6

SCALE 1/2

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NO. OF SHEETS PER SET | SEE SHEET 1 | SHEET

APP'D | CHK | DRAWN

ISSUE AUG 26 1952

ture are being considered. If scaled to 55,000 mc, the slotted sheet would be .002" thick and the preparation of such a sheet, in molybdenum, by photographic etching is being attempted. If the etching is successful, a tube design that has been considered to go with the sheet will be attempted. The longitudinal gap in the middle of the sheet is a means of preventing buckling in case of heating up, and may also serve as a place to put a tightly-focused round beam.

8.222 Oscillators ("Backward-Wave")

A demountable oscillator tube based on the circuit arrangement of Model III has been designed and construction is under way. The part which forms the ridged guide is to be wound with gold-plated molybdenum wires or tapes and sintered. The tapered coupling waveguide at the gun end is being electroformed. At the opposite end, the circuit is to be terminated with wedges of graphite-loaded ceramic. A gun to produce a ribbon beam in a strong magnetic field with about 1 A/cm^2 current density has been designed and is being constructed.

8.23 Other Slow-Wave Structures

Preliminary investigations were made on other forms of slow-wave structures, which are illustrated in Figs. 7 and 8.

The structures identified as Models 6 and 6A in Fig. 8 are very similar to those described above, the grid of wires or tapes being transferred from the top wall of the guide to the top of an open ridge inside the guide. Models 6 and 6A differ

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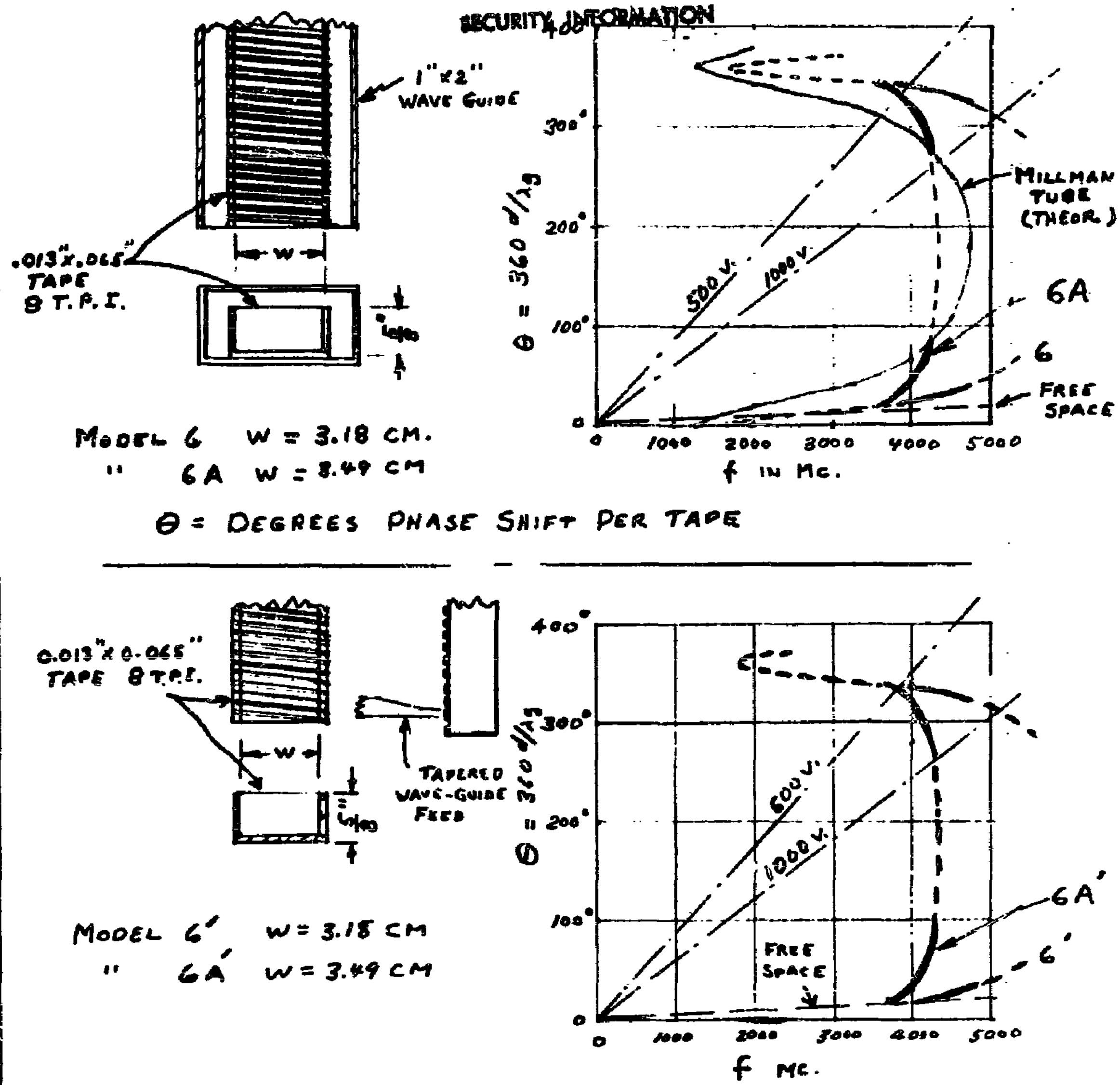

FIG. 7

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S. D. ROBERTSON

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SLOW-WAVE CIRCUITS FOR SPATIAL HARMONIC OSCILLATORS

SCALE $\frac{1}{2}$" = 1"

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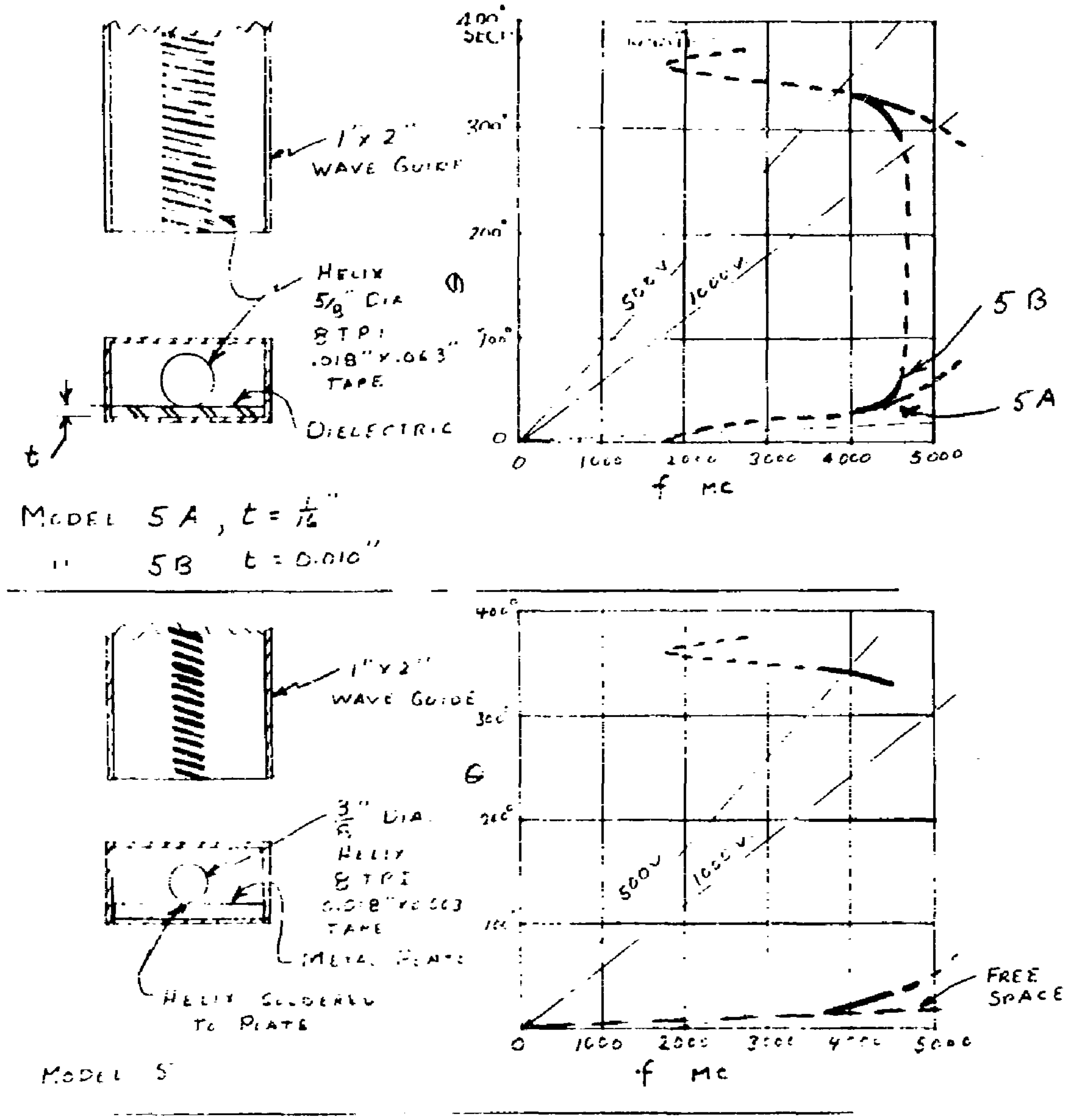

FIG 8

SCALE 1/2" = 1"

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only in the width of the ridge, and their phase characteristics are plotted at the right. (Since the measuring equipment was operable over only a limited bandwidth, it was not possible to measure the phase over the entire propagating ranges of the models. The dotted portions of the curves therefore represent an estimate of the shape of the curves in the unmeasured regions.)

Models 6 and 6A offer the possible advantage that the waveguide itself may serve as the vacuum envelope.

It was found that, if the outer guides of Models 6 and 6A were removed and the wound ridges alone excited at one end by means of a tapered waveguide, they too behaved as slow-wave circuits. They have been identified as Models 6' and 6A', and their phase characteristics are shown.

All of the above wound structures require a flat electron beam which just grazes the edges of the wires or tapes. It would be preferable in some cases to use a Brillouin focused beam of circular cross section where a lower magnetic field would suffice. The structures shown in Fig. 8 were conceived to be used with circular beams. Models 5A and 5B make use of a helix which is separated from the bottom wall of the waveguide by a thin sheet of dielectric. The helix is therefore loaded by the capacitance to the guide wall. Model 5 and 5A differ in the thickness of the dielectric. In Model 5 (lower figure) a somewhat smaller helix is soldered to the bottom of the guide.

An important advantage of circuits of the last two types is that the helix and beam diameters would be several times larger than in the conventional traveling-wave tube.

8.3 Millimeter-Wave Measuring Equipment

A single-detection, millimeter-wave measuring set has been assembled and put in operation. Parts are being made for a phase modulator of the type developed by D. H. Ring, which will extend the dynamic range of the equipment. This measuring equipment will be used for studying the operating characteristics of the helix amplifier as well as other tubes now being developed.

8.4 Miscellaneous

An additional demountable vacuum pump station, complete with a cold trap, is now being assembled.

9. Spatial Harmonic Amplifier (C. F. Quate and R. M. Rogers)

During this period several models of the 5.4 mm scaled version of the Millman tube have been assembled and pumped. Successful operation has not as yet been obtained however because of our inability to maintain cathode activity in this tube. The activity trouble has been attributed to the poisoning effects of the fine wire control grid just off the cathode surface. Also in this tube the collector design is such that it cannot be separately outgassed.

At present a new gun design is being incorporated into the tube and also an external collector is being added. In addition it has been found that the vacuum windows are sufficiently far from the beam so that kovar at this point does not disturb the beam transmission. Therefore, kovar is being used on the windows and lead in seals. This will allow the main body of the tube to be outgassed in the 600° region.

For future tubes some thought is being given to ceramic seals so that the entire tube can be baked out above 600°C. The primary problem involved is that of ceramic windows across the waveguide, and we are currently testing several schemes such as flat discs slanted across the guide.

10. Millimeter Wave Low Voltage Reflex Klystron (E. D. Reed, L. B. Luckner, F. P. Drechsler)

10.1 General

The principal effort in the past quarter was directed towards the detailed mechanical design of the 5.4 mm reflex klystron. This phase of the work has been essentially completed and is described below in detail. Other phases which received attention were the electron gun, the output window and the cavity. Work was also continued on the demountable 6.25 mm fixed tuned reflex klystron.

10.2 Final Design of 5.4 mm Reflex Klystron

The layout of the 5.4 mm reflex klystron, selected from a number of alternative designs, is shown in Fig. 1. Referring to this figure, cavity block, 1, is seen to be identical with that shown by Fig. 3D in the last report (June 1952). The cavity, inner and outer choke sections as well as the output wave

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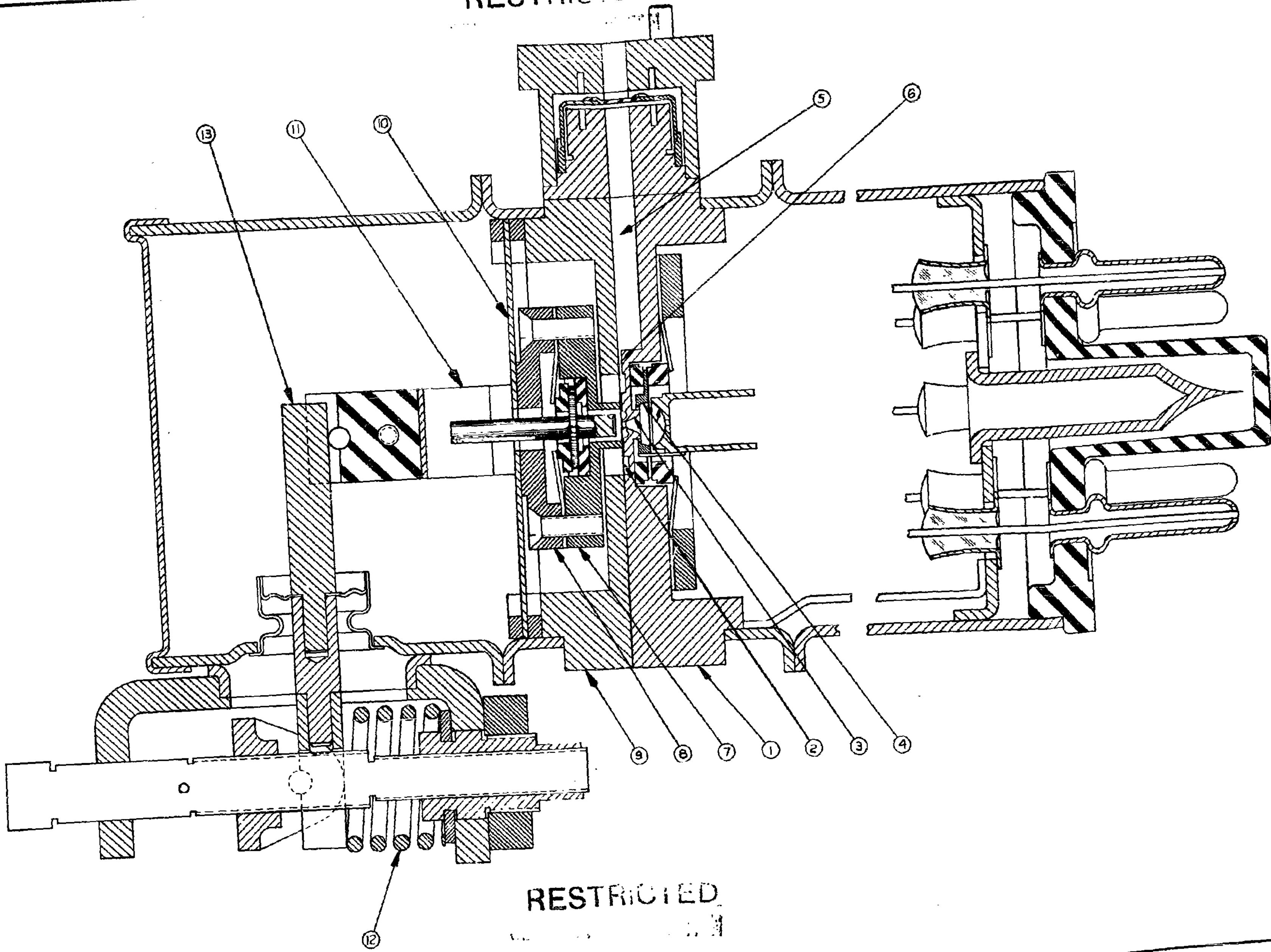

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FIG. 1

guide are hubbed out of a solid copper blank with a number of machining operations required to provide a flat surface for the copper diaphragm, 2, and a properly contoured focusing anode, 3, for the electron gun, 4. Coupling between the output waveguide, 5, and the outer choke section of the cavity is obtained by means of the quarter wave transformer, 6, the height of which determines the tightness of coupling. This height, since it might be subject to change in early experimental tubes, is not obtained from the original hubbing but is milled into the cavity block upon completion of the other machining operations. The repeller housing, 7, and the upper cavity block, 9, are gold brazed to the diaphragm to form a separate subassembly. The latter is BT brazed to cavity block after the G1 and G2 aperatures have been optically aligned. Microscope techniques are also employed in aligning both repeller and electron gun, since it was found that the required accuracy could not be obtained by self-alignment. Spacings are maintained through the use of ceramic washers, selected to compensate for parts tolerances, and the electrodes are clamped in place firmly by disc springs. The initial gap spacing is adjusted for cold resonance beyond the upper frequency limit of the tuning range by means of the stiff restoring spring, 10, and tuning motion is transmitted to the diaphragm through the rigid link consisting of details 7, 8 and 11. The actuating device is the side-arm tuner, 12, similar in construction to the one used in the Western Electric reflex klystrons 419A and 431A, where it has resulted in tubes having a high degree of frequency stability. For this application it was modified such that one revolution of the differential tuning screw produces a diaphragm motion of about 0.8 mil. A noteworthy feature of this mechanism provides for a non-rigid connection between tuning arm, 13, and link detail, 11. Thus undesired changes in grid spacing due to differential expansion in pumping, bake-out and initial warm-up are avoided. Contact between the tuning arm and the insulated pin in tuning link, 11, completes an external alarm circuit, thereby indicating the point at which the initial gap is closed and actual diaphragm motion commences.

The tube can be fully assembled in mount form and all connections made to the stem before the vacuum envelope is completed.

Present plans call for a glass output window similar to the one used in the Millman tube, but experimental work is also in progress on a mechanically simpler mica window. Either of these windows will be attached to the output waveguide by brazing.

Major points of difference between the above design and the earlier one described in the first report of Sept. 1951 on pp. 50-54 are these:

(a) Cavity, choke section, coupling circuit and output waveguide are hubbed out of solid copper, thus providing better electrical conductivity and cooling.

(b) Diaphragm has been changed to flat copper sheet.

(c) Cavity contour has been rounded in order to provide smooth surface finish and improved shape factor.

(d) There are no brazed joints in primary cavity.

(e) Cavity block forms part of tube envelope, hence, if necessary, making efficient forced-air cooling possible.

(f) Length of waveguide run between cavity and output window has been reduced by a factor of two.

(g) Electron gun, repeller and G2 aperture are aligned with G1 aperture by optical means.

(h) Positioning of gun and repeller in direction parallel to axis of tube is obtained by ceramic spacers.

(i) Output waveguide is coupled to outer choke section via single-step quarter-wave transformer, resulting in mechanically simple configuration and good broad-band properties.

(j) Thermal tuner has been replaced by mechanical tuner.

10.3 Cavity Testers

Preliminary hubbing tools were prepared and a number of successful impressions made proving the feasibility of producing the cavity block by a combination of hubbing and machining. One such impression was used in the tunable cavity tester shown in Fig. 2, operating in the 5.4 mm band. This tester could be tuned from 5-6 mm, and thereby confirmed the cavity dimensions which had been obtained by scaling from 4000 mc brass models.

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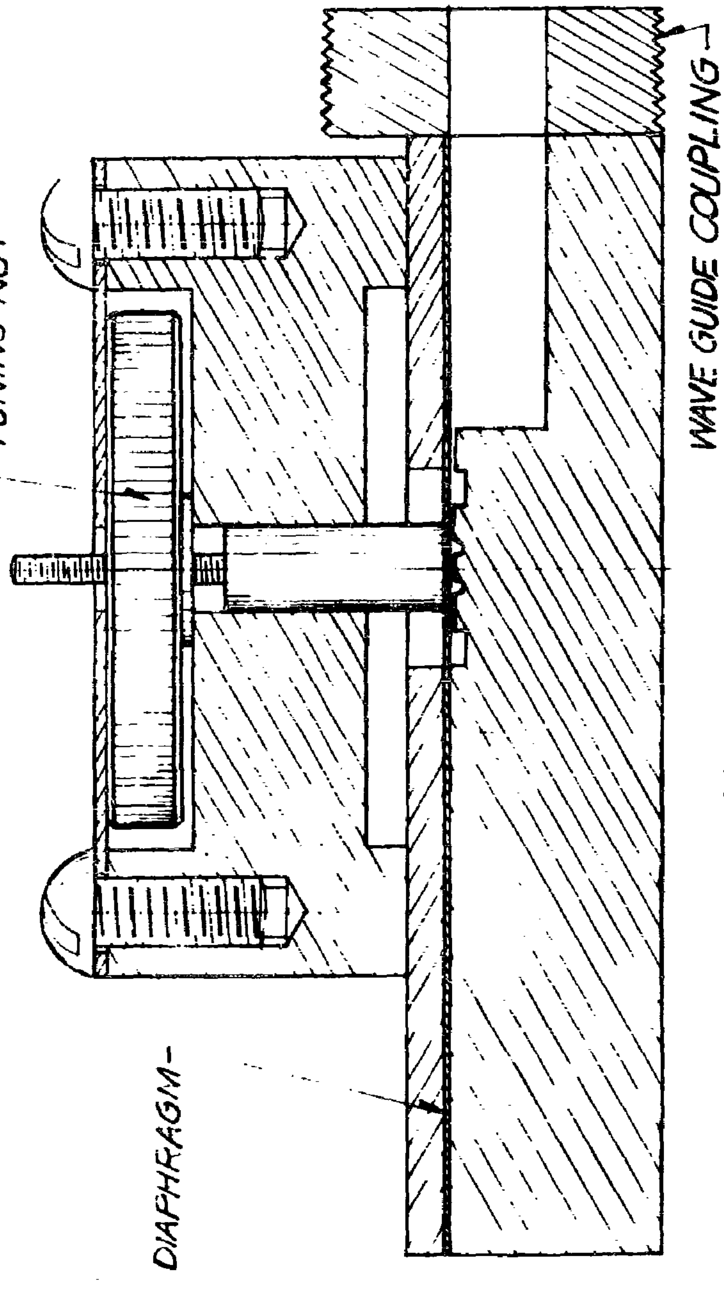

FIG. 2

P-		
ENGR		
DRAWN		
ISSUE		

USED ON	DWG

SCALE

WESTERN ELECTRIC CO. INC
ENGINEER OF MANUFACTURE
CENTRAL OFFICE EQUIPMENT

BELL TELEPHONE LABORATORIES, INC

MATERIAL	FINISH

DIMENSIONS UP TO AND INCLUDING 72 INCHES EXPRESSED IN INCHES. NON-LIMITED DIMENSIONS, OTHER THAN SIZE OF RAW MATERIAL, SHALL BE HELD WITHIN FRACTIONAL ___ DECIMAL ___

DWG SIZE 1S

P-

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10.4 Present Status

At the present time the design and detailing of the 5.4 mm tube have been completed with some work yet to be done on assembly tools and jigs. Parts have been ordered and work is in progress on the final, polished hubbing tool.

The mechanical stability of the electron gun design is being investigated by life tests, in which the parts are cycled between room - and operating temperature by the application of heater power.

Having obtained mica-to-metal vacuum seals on preliminary testers, a final design of mica window for this tube is now being worked out.

11. The Possibility of Using Dielectrics as Slow-Wave Structures for Traveling-Wave Tubes in the Millimeter and Sub-Millimeter Ranges (H. Suhl)

One of the obstacles to the extension of traveling-wave tube techniques to shorter and shorter wavelengths is the manufacture of a sufficiently accurate slow-wave circuit. This particular obstacle could be overcome if a natural circuit could be found to replace the man-made structure. In principle, a slab of material having either a very high permeability or a very high dielectric constant provides just such a slow-wave circuit. A strip of the substance of thickness between one half and one full length of the electromagnetic waves in the unbounded medium and deposited on a metal backing will propagate waves along it with a speed only a little above the speed on the unbounded medium. The electromagnetic field will extend some distance out from the strip and will interact with an electron beam running close to it, or at grazing incidence.

An analysis of this proposal has been made, and some of it is found to run very close to usual traveling-wave tube calculations. Some of the criteria for a good traveling-wave tube therefore apply directly. Thus it is still true that to obtain a high gain, the cold circuit should have a high impedance. Now the impedance of such a slab is of order $\sqrt{\frac{\mu}{\epsilon}}$. A high dielectric constant therefore means a low impedance and therefore little gain per wavelength. This is compensated somewhat by the fact that the structure can be many wavelengths long in the frequency range of interest. But it is not certain that coherence of the field can be maintained over a great many wavelengths in view of possible non-uniformity in the material. A high permeability on the other hand gives a high circuit impedance. But few if any materials are known which give a high permeability in the millimeter band without the use of a fixed auxiliary magnetic field. Such a field can produce a gyro-resonance of the electron spins and thus a high permeability in a limited frequency range.* That range is close to the cyclotron frequency associated with the magnetic field. Hence

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*C. L. Hogan, B.S.T.J. January 1952.

at 6 mm a magnetic field greater than 10 kilooersteds is required. For wavelengths shorther than 6 mm, the necessary apparatus would thus be very cumbersome.

It is also true, for that matter, that dielectrics with a high ε are very rare in that frequency range. But there exists the possibility of using a material with strong absorption lines due to lattice vibrations. Near an absorption line the dielectric constant rises sharply. If the line is sharp enough, the accompanying loss may not be excessive.

As an example we consider one of the few materials for which sufficiently extensive data are available: rock salt, whose absorption line near 60 microns has been examined.* From these data it appears that the refractive index ought to reach 10. Neglecting losses, the gain should be

.0248 db/circuit wavelength

at a beam current density of 30 amps/cm^2, and with a slab one half a circuit wavelength (= 3 microns) thick. Losses will reduce this figure by a factor of about 3, giving

.0083 db/circuit wavelength.

This figure is very low compared with that for a conventional tube, where the gain is around 2 db/circuit wavelength. Thus to get a net gain of 20 db, 2400 circuit wavelength, or a total length of 1.42 cms, would be necessary. This does not seem a great length when stated in rms, but it is hard to say whether

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*Cartwright & Czerney, Zeitschrift für Physik, Vol. 85, p. 269, 1933.

traveling-wave tube action can be maintained in practice over 2400 circuit wavelengths.

Absorption lines at longer wavelengths would give a greater gain. The scale factor is (wavelength)$^{2/3}$. Thus for 600 microns (.06 mm) the gain would be about .036 db/circuit wavelength. At 6000 microns (.6 mm) it would be .178 db/circuit wavelength, and at 6 mm it would be .83 db/circuit wavelength, it being assumed that the refractive index is ten, the current density 30 amps/cm^2, and the slab thickness one half a circuit wavelength. (There can be little doubt that materials with absorption lines up to at least .6 mm do exist.)

The principal handicap of a tube of this kind would be the low circuit impedance. If this could be raised substantially, more gain could be expected. An interesting possibility in this direction lies in the use of "dichroic" media. These are anisotropic materials with absorption lines whose frequencies depend on the polarization of the wave. With such a material it is possible to work a frequency at which the dielectric constant for one polarization is high ($= \varepsilon_1$, say) while that for another polarization is actually negative ($= -\varepsilon_2$, say). If the strip is cut so that the polarization direction for ε_1 is normal to it, while that for ε_2 is along it, it turns out that the propagation speed is now <u>below</u> the speed in an infinite medium with isotropic dielectric constant ε. Hence the beam velocity can be decreased (by an amount $\sqrt{|\varepsilon_2|/\varepsilon_1}$ for the fastest mode), and for the same current density, the charge density can therefore be increased. Combining this fact with the apparently greater circuit impedance, an increase in gain by a factor

$$\left(\frac{4}{3}\right)^{1/4}\left(\frac{\varepsilon_1}{|\varepsilon_2|}\right)^{5/6}$$

becomes possible. If $\varepsilon_1 = 100$ and $|\varepsilon_2|$ is between 2 and 4, this factor is between 9 and 25. Thus at .6 mm, a gain between 1.6 and 4.45 db/circuit wavelength should become possible.

The difficulty here lies in an almost complete absence of detailed knowledge on absorption lines in the very far infrared. This, in turn, is caused by a lack of sources of sufficiently high power infrared radiation. We have to conclude therefore that successful application of the principles outlined here requires detailed information which is not available at the present time.

12. Extradigitron (D. Hagelbarger and L. R. Walker)

The 60-wire tube mentioned in the last report has been built and tested with disappointing results. Its performance was decidedly inferior to the earlier 30-wire tube. The tube had a number of mechanical difficulties and a spurious mode in the output window circuit. Rather than attempt to rebuild it a tube with 45 wires is being designed.

Techniques

Some experiments in making extradigitron circuits for 5.4 mm have been made. By annealing the wire during winding and winding on a square polished mandrel we have obtained helices where the 0.0160" spacing varied less than ± 0.0002".

13. Backward Wave Oscillator (D. Hagelbarger and L. R. Walker)

Further measurements on a 1000 mc model extradigitron circuit confirm our observation that the characteristic impedance of the modes used for a backward wave oscillator changes rapidly with frequency. A factor of 3.5 in level over a 10% bandwidth in the region where the circuit is most likely to be used was measured. This means that a broad-band match cannot be obtained with a smooth taper since the output waveguide impedance is relatively constant over the same bandwidth. We have however obtained matches a few per cent wide by simple means.

An analysis has been made of the starting currents to be expected in a characteristic backward wave oscillator. These seem to agree with experiments made at Stanford. It is clear from the analysis that backward wave oscillation is an interference phenomenon not necessarily involving amplified individual waves.

14. Propagation in Waveguides Containing Ferrites (H. Suhl and L. R. Walker

Calculations extending the results obtained in the Physical Review on propagation in cylindrical waveguides filled with a material exhibiting Faraday rotation are being carried on. A portion of this work is being carried out using Bell System funds as a part of a Waveguide Research Program.

15. High Frequency Measurements (C. T. Goddard and N. C. Wittwer)

15.1 Model Studies

A preliminary study of the limitation in high frequency performance imposed by the use of conventional button stems was presented in the last quarterly report. Since the input admittance characteristics are controlled by passive feedback parameters wherever they may occur, it was recognized that a study of cathode lead inductance alone is not sufficient. A complete study must include the mutual inductance effects of all leads and any capacitive elements which contribute to feedback. The scale model approach is appealing in its simplicity and in the ease with which modifications may be checked for improvement. For this reason a large scale model complete in all details has been constructed for test. The model mounted on a 5" diameter simulated stem is pictured in Fig. 2.1. The circuitry on the small chassis has been designed for input conductance measurements by a substitution method.

The model and test assembly were completed in the recent past.

15.2 Direct Measurements

From measurements of short circuit admittances made on conventional tubes several years ago it is known that similar measurements on the newer high transconductance types present new problems. As operating circuit impedance levels become lower the range of interest in input loading conductance in-

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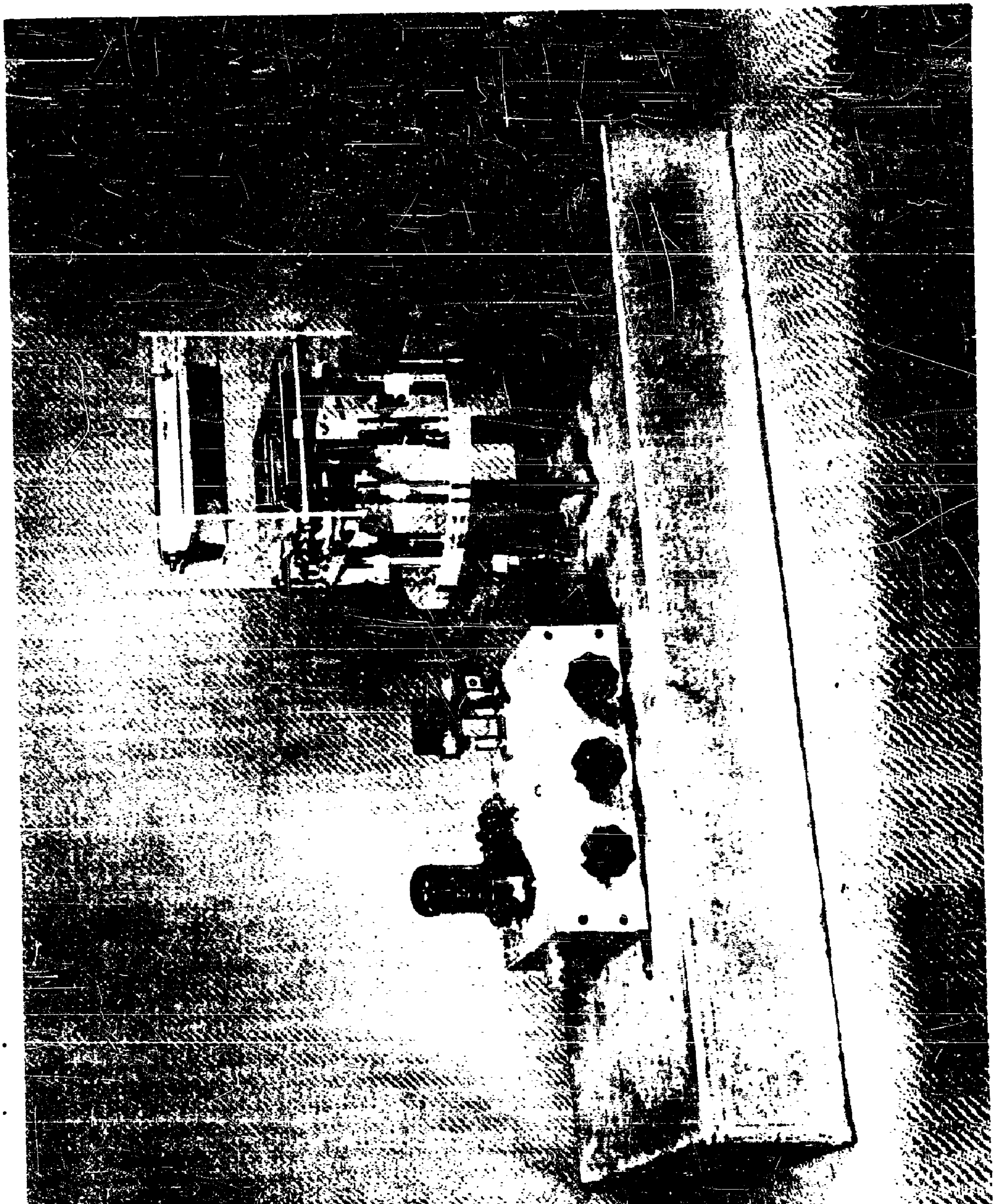

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creases such that Q-meter techniques become inaccurate. At the other extreme in measuring techniques, transmission line methods are well suited to very high conductances. Unfortunately the range of conductance values of interest in the M1686 type of tube is the order of 500 to 5000 micromhos where neither of the above methods is particularly reliable.

In the range 50 to 400 megacycles it appeared that a modified transmission line technique held considerable promise. The General Radio type 1602A Admittance Meter appeared most suitable for this adaptation. By selecting a new standard or reference termination having a conductance value approximately centered in the range of conductances to be measured, the accuracy of measurement was considerably improved. A trial run using the Western Electric type 404A vacuum tube as a test sample yielded a frequency-squared dependence of short circuit input conductance as the theory predicts. Absolute values of measured conductance were of the same order of magnitude as calculations based on estimated cathode lead inductance.

A few finishing touches and a carefully designed test socket will be required before use on the M1686 type structure is feasible. The equipment is intended as a direct high frequency check of the studies being conducted by scale model techniques.

As a byproduct of these studies it has been found that input resonance effects due to the inductance of the grid lead may present as big a problem in tube design as the feedback inductance of the cathode lead. Series resonance with the input capacitance was encountered in the range 170 to 250 megacycles with the large M1686 tube - the exact value depending on the care with which the inductance of the input connection was minimized. Operation of tubes considerably below this resonant point will still present difficulty in interstage network design. Further considerations of these matters are reported in the next section.

16. Problems Associated with Broad-Band I.F. Amplifier Tubes (C. T. Goddard and N. C. Wittwer)

16.1 Impedance Transformation, Bandwidth and Operating Frequency

The figure of merit for a broad-band amplifier tube (representing the product of voltage gain and bandwidth) is,

$$F = K_o \frac{Gm}{\sqrt{C_{in}\, C_{out}}} \tag{16A}$$

The constant "K_o" is dependent on such factors as the complexity of interstage networks used, the method of specifying bandwidth and flatness and whether the circuits are matched. This common equation may be rewritten to emphasize the problems of tube and circuit designer as follows:

$$F = K \frac{Gm}{C_{in}} \sqrt{\frac{C_{in}}{C_{out}}} \tag{16B}$$

The term Gm/C_{in} is primarily a tube design parameter and is a function only of the input geometry and operating current density for the tube.

The remaining term in equation (16B) is a parameter of interest to both the circuit and the tube designer. To the circuit man the term $\sqrt{\frac{C_{in}}{C_{out}}}$ represents the square root of the impedance transformation which must be obtained in the inter-stage network for optimum bandwidth performance. In the design of tetrodes for broad-band operation in Bell System applications and under this contract every effort has been made to reduce the output capacitance to a minimum. The resulting large ratio of input to output capacitance represents a real improvement in tube design but presents a new problem to the circuit man - the problem of high ratios of impedance transformation over very broad bands.

In any four-terminal network of fixed complexity in design there is an octave-wise limitation to the frequency band over which a given impedance transformation may be easily obtained. A quarter wave transmission line, for example, may be used to couple two different resistive impedances over a large fraction of an octave if the ratio of impedances is small. As the ratio is increased, the bandwidth of the device becomes more restrictive until in the limiting case it approaches single frequency operation. The same is true for tuned coupled circuits, for here the coefficient of coupling must be held constant for a desired fractional bandwidth. As the impedance ratio is raised one coil of the coupling pair becomes correspondingly smaller until it may be difficult or impossible to attain the required coupling coefficient.

These problems may be circumvented by the use of more complex interstage networks such as double-step quarter wave transformers or synchronously tuned triplets. The added complexity in design, alignment and servicing makes this approach rather unattractive however.

The more direct approach of designing for higher center frequency of operation has many advantages. Not only does the fractional band required for obtaining a desired bandwidth become smaller, but the system design requirements are simplified. Broad-band high resolutions systems are usually characterized by operation at frequencies higher than 10,000 megacycles. As the system frequency is raised for higher resolution, broader bands are required and a "low" frequency I.F. amplifier aggravates the problem of carrier frequency filters for image suppression. Thus from a system point of view it is desirable to consider doubling the I.F. frequency every time the carrier frequency is doubled.

Based on the above considerations, one goal of the work under this contract has been the investigation of factors which limit the top operating frequency for space charge control electron tubes in conventional or semi-conventional stem-and-bulb enclosures.

In the next section the factors limiting the percentage bandwidth attainable with one design of popular interstage network is evaluated.

16.2 Limitations of "T Equivalent" Synchronously Tuned Double-Tuned Transformers

In Fig. 3.1 the transmission characteristic of a particular type of interstage network is plotted. The design parameters for this case were evaluated several years ago for Bell System applications. The coupling coefficient as defined provides a slightly "double-humped" characteristic with a transmission loss at mid-band of the order of perhaps a few hundredths of a db. The data have been replotted here to demonstrate the problems discussed in the previous section.

The "T" equivalent shown has been found particularly attractive because the inductive elements may be preset by conventional "Q-meter" techniques. There is no magnetic coupling between elements.

As the required impedance transformation for this network is increased the maximum coupling coefficient attainable (determined by $1 - K\sqrt{\frac{C_2}{C_1}} = 0$) decreases. The abscissa (plotted as fractional bandwidth for several values of the coupling coefficient) shows that the total bandwidth decreases on a percentage basis as the impedance transformation ratio is raised. Thus for tubes having a very high figure of merit there is a minimum frequency at which this circuit may be used for broad-band amplifiers.

From the data of Fig. 3.1, Fig. 3.2 has been plotted. Here the curves A, B, C and D determine the maximum bandwidth

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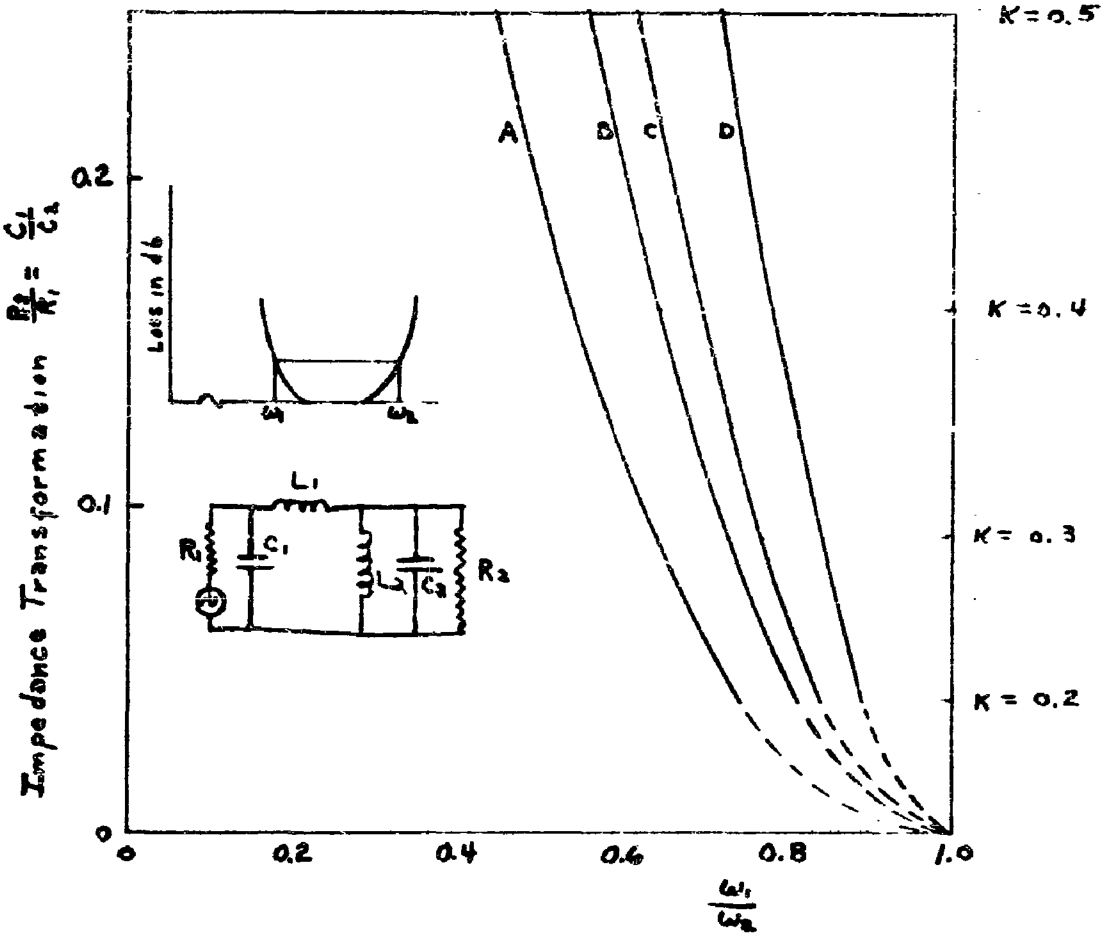

Circuit design in accordance with figure 3.1

Curve A : 3db band width
B : 1db band width
C : 0.5db band width
D : 0.1db band width

Impedance Transformation vs Fractional Frequency Range

Fig 3.2

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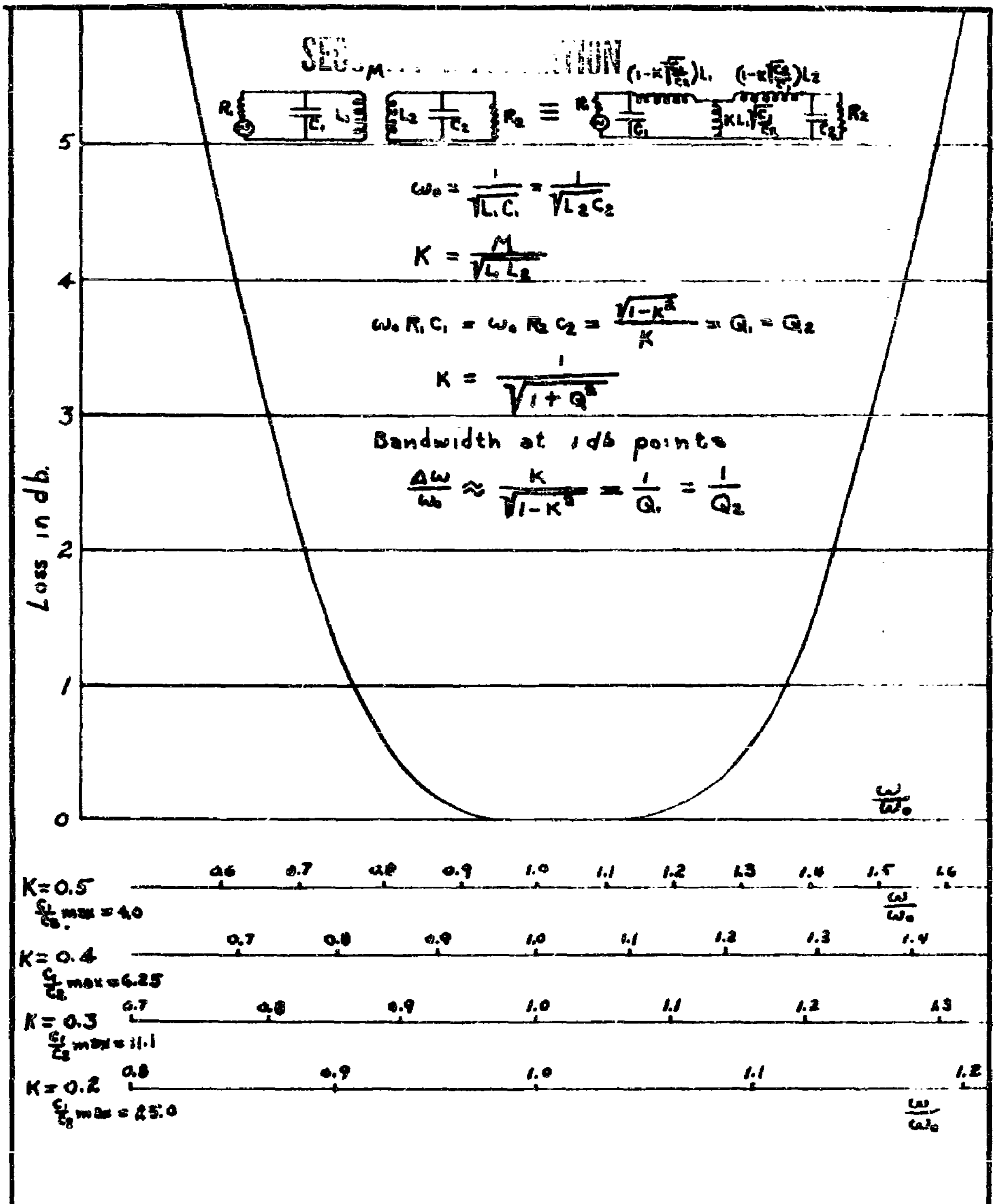

Double Tuned Transformer

8/29/52

Data obtained in part from SL327430 (V.C. Rideout, 8-12-46)

Fig 3.1

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(on an octave basis) over which a desired impedance transformation may be attained. The curves cover bandwidths measured to the 0.1, 0.5, 1 and 3 db points to permit a simple solution for multi-stage amplifier requirements.

16.3 Application to the M1686 Tetrode

In order to determine the optimum size of tube in terms of cathode area for the M1686 electron tube design, the methods of the previous two sections have been applied in part. A final solution must await the conclusion of direct and simulated high frequency measurements now in progress.

Further estimates of performance characteristics for the M1686 with the original cathode area of 0.7 square centimeters and of a one half and one quarter size tube are given in table 3.3. For comparative purposes similar figures for the Western Electric 436A of recent design are given. It is seen that the highest figure of merit is obtained with the tube of largest cathode area. This is a natural result of the minor effect which stray circuit capacitance has on the ratio C_{in}/C_{out}. However, the large impedance ratio of approximately 10:1 would require operation at an impossibly high frequency for such a large tube using the circuitry of Figs. 3.1 and 3.2.

From Fig. 3.2 and table 3.3 we may estimate the operating frequency required and the 1.0 db bandwidth attainable for a 10 db per stage amplifier using the M1686 model estimates of table 3.3. These are:

Tube Model	f_o (mc)	Δf (mc)
M1686	450	160
$\frac{M1686}{2}$	240	100
$\frac{M1686}{4}$	160	80

Obviously the M1686 performance is impossible, since the input circuit is resonant considerably below 450 megacycles. The one half size tube does not appear impractical, and the one quarter size tube appears very reasonable indeed.

The above figures are rough estimates at best. A more thorough analysis must await the experimental determination of input loading and effective grid-lead inductance.

Estimates of Characteristics Attainable with Tubes in Operating Circuitry

Table 3.3

	ML686	ML686 2	ML686 4	436A
Cathode Area (cm^2)	0.7	0.35	0.18	0.66
C_{g_1-k} (hot)	24	12	6	12.8
$C_{g_1-g_2}$	7	4	2.5	4.6
C_p' (internal)	0.6	0.3	0.15	
C_p (in enclosure)	1.2	0.9	0.75	3.3
g_1-g_2 spacing (inches)	.006	.006	.006	.012
Cathode Current (mA)	36	18	9	33
G_m (micromhos)	60,000	30,000	15,000	28,000
C_{in} (note 1)	30.5	18.0	10.5	21.9
C_{out} (note 2)	3.2	2.9	2.75	5.3
F (note 3)	965	661	445	390

Note 1. $C_{in} = C_{g_1-k} + C_{g_1-g_2} + 2.0$ (circuit strays)

Note 2. $C_{out} = C_p + 2.0$ (circuit strays)

Note 3. $$F = \frac{G_m}{2\pi \sqrt{C_{in} C_{out}}}$$

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